Final Technical Documentary Report

DEVELOPMENT OF INTEGRATED MICROWAVE COMPONENTS FOR GROUND-BASED PHASED ARRAYS (ELECTRONIC STEERING MODULE)

January 1968

Lincoln Laboratories Project No. C-595, dated 17 April 1967
USAF Contract AF 19(628)-5167

1 of 1

By

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Lafe R. Pfieffer, Jr.

Prepared for
Lincoln Laboratories
Massachusetts Institute of Technology
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Texas Instruments Incorporated
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ABSTRACT

This is the final technical documentary report for the Electronic Steering Module program, Contract No. C-595, prepared for Lincoln Laboratories, Massachusetts Institute of Technology. This report documents all efforts expended from contractural start date, 17 April 1967, to 15 January 1968. The overall goal of this program was to design an element steering module that would be easily produced in large quantities at a minimum cost. The effort of this program was directed toward the design of the ESM module in a microstrip form utilizing thin-film techniques to form the resistors, capacitors and transmission-time elements on a ceramic substrate.

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FOREWORD

Program Director:	Mr. Carl Blake Lincoln Laboratory Massachusetts Institute of Technology
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LIST OF ILLUSTRATIONS

Report No. 03-68-03

Texas Instruments Incorporated
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Microwave Products Department

January 1968

Final Report
Electronic Steering Module

Reference: Massachusetts Institute of Technology, Lincoln Laboratory, P.O. Box 73, Lexington, Massachusetts 02173, Purchase Order No. C595 dated 4/17/67.

Acknowledgement: USAF Contract AF19(628)-5167

SECTION I

INTRODUCTION

This is the final technical documentary report for the Element Steering Module program, Contract No. C-595, prepared for Lincoln Laboratories, Massachusetts Institute of Technology. This report documents effort expended from contractual start date, 17 April 1967, to 15 January 1968.

The overall goal for the element steering module development program was to produce a module design that was easily produced in large quantities at a minimum cost. The large quantity, low-cost goal is necessary for the use of this type module in large phased-array defense-type radars. The effort of this program was directed toward the design of the ESM module in a microstrip form utilizing thin-film techniques to form the resistors, capacitors and transmission line elements on a ceramic substrate. The mixing and switching diodes in a beam-lead form are attached to the thin-film circuit by thermal-compression bonding or another similar process.

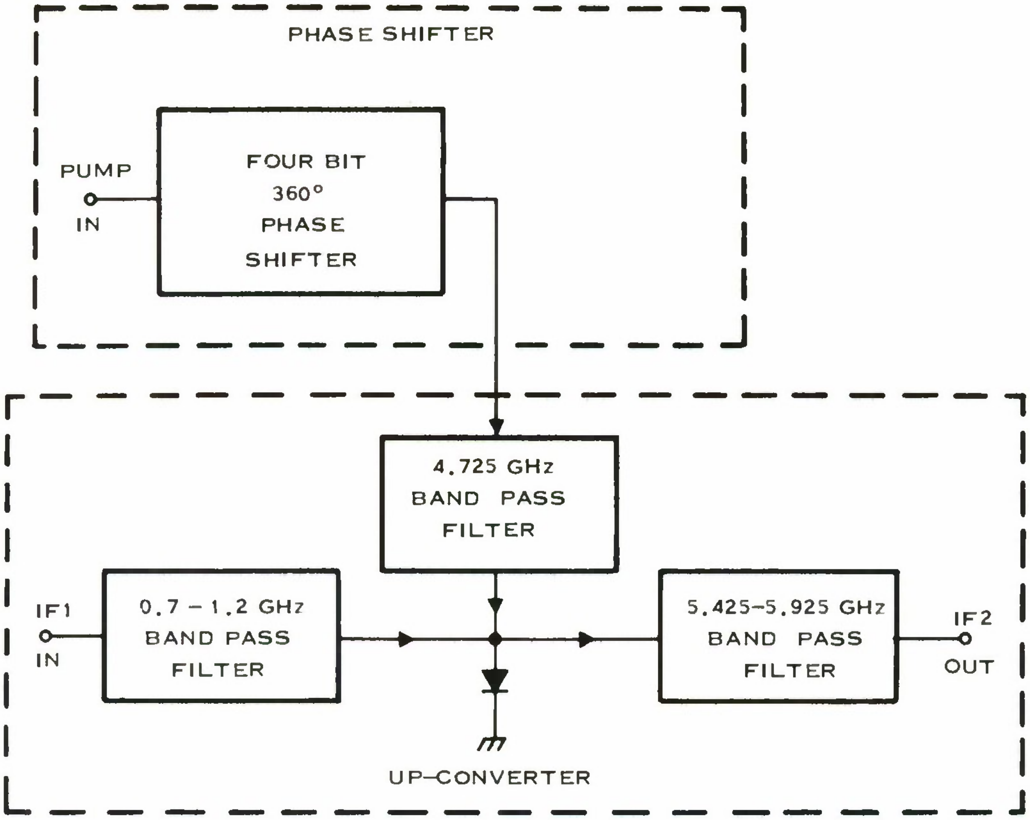
This hybrid approach is the most economical form for microwave circuits, such as the ESM module, involving filter structures in the 1 to 10-GHz range that require a large area for implementation. The thin-film approach, using photo-etching techniques, is also capable of high-reproducibility which allows the circuits to be mass produced without the necessity for tuning adjustments.

SECTION II
DESCRIPTION OF SYSTEM

A block diagram and parameters for the proposed element steering module are shown in Figure 1. It is divided into two basic, functional sections, a phase shifter and a frequency translator (or up-convertor). The purpose of the ESM module is to provide beam steering and frequency translation for a large, phased-array receiving system. The beam steering is accomplished by adjusting the phase of a 4.725 MHz pump signal for a variable reactance up-convertor. The input to the up-convertor is a 500 MHz band from 0.7 to 1.2 MHz converted to the upper side-band from 5.425 to 5.925 MHz.

The phase is shifted in 22.5° increments using a 4-bit binary scheme. This is accomplished by using diodes to switch the values of a shunt reactance across a transmission line to produce incremental phase shifts of 22.5°, 45°, 90°, and 180°. The advantages of this type phase shifter are: small in size, light in weight and, it is easily controlled by standard low-level integrated-logic circuits.

The up-convertor was proposed rather than a conventional mixer due to a 3-dB gain that can be realized rather than the approximately 8-dB loss resulting in the use of a mixer. The up-convertor system results in a net 11-dB system gain. This up-convertor circuit consists of a low-pass filter at the input, a band-pass filter at the output, and a varactor diode with the pump-tuning circuitry connected in series between the input and output filters. The varactor diode is the step-elastance type and was chosen rather than the usual variable-capacitance type in order to reduce the conversion-gain variations occuring with pump-power variations using a variable-capacitance diode.



ELEMENT STEERING MODULE SPECIFICATIONS

INPUT FREQUENCY	700 – 1200 MHz
OUTPUT FREQUENCY	5.425 – 5.925 GHz
PUMP FREQUENCY	4.725 GHz
PUMP PHASE SHIFTER	4 BITS
PHASE TRACKING ACCURACY	$\Delta\Phi_{max} = 6°$ rms FROM AVG.
PHASE SHIFTER SWITCHING TIME	3 μs
PUMP POWER	50 mW
LOWER SIDEBAND ISOLATION	20 dB
PUMP ISOLATION	20 dB
CONVERSION GAIN	>0 dB
AMPLITUDE TRACKING	$\pm$1 dB
INPUT/OUTPUT IMPEDANCE	50 Ω
INPUT/OUTPUT VSWR	1.5
INPUT LEVEL FOR 1 dB OF COMPRESSION	0 dBm

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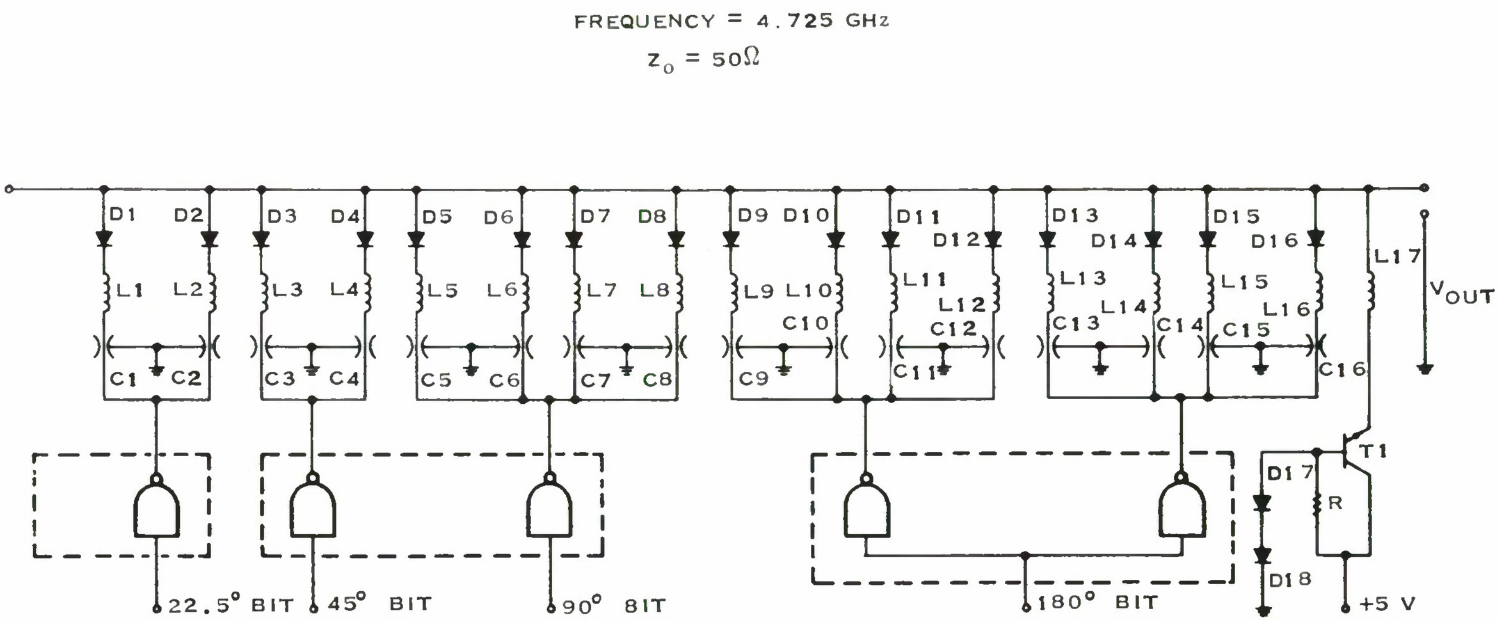
Figure 1. Block Diagram and Parameters for Element Steering Module (ESM)

SECTION III
CIRCUIT DEVELOPMENT

A. PHASE SHIFTER

The schematic diagram of the proposed phase-shift network is shown in Figure 2. This is a diode-loaded line-type phase-shift network and has been discussed by J. F. White.[1/] This is an iterative-type circuit, having a canonical form consisting of a length of transmission line symmetrically loaded at its ends by small susceptances whose values are switched by means of diodes. The value of the susceptances, the line impedance, and the line length are chosen to obtain the desired phase shift while, at the same time, maintaining a low VSWR by mutually cancelling the reflections produced by the susceptances. This type of phase shifter is capable of handling high power by designing each section for small phase shifts (<22.5°) which in effect decouples the diode from the main transmission line. It can be shown that this also allows operation over a wider bandwidth for a given phase error.

The disadvantage of designing for a low phase shift per section is the larger number of sections required for a given total phase shift. The proposed phase-shift network shown in Figure 2 uses a 22.5°-section and seven 45°-sections to obtain a 4-bit (or 16-step) 360°-phase shift. The 90° and 180°-bits were realized by combining the appropriate number of 45°-sections. In order to reduce the number of sections, and hence the size and cost, a study was made of realizing 90° and 180°-bits with other than 45°-sections. The only logical possibility is a 90°-section, using one for the 90°-bit and two for the 180°-bit.



COMPONENT LIST			
D1, D2	C = 0.067 pF	C1–C16	C = 75 pF
D3–D16	C = 0.13 pF	R	R = 150Ω
L1, L2	L = 8.4 nH	T1	2N2222
L3, L16	L = 4.4 nH	D17, D18	REFERENCE DIODE
L17	CHOKE		

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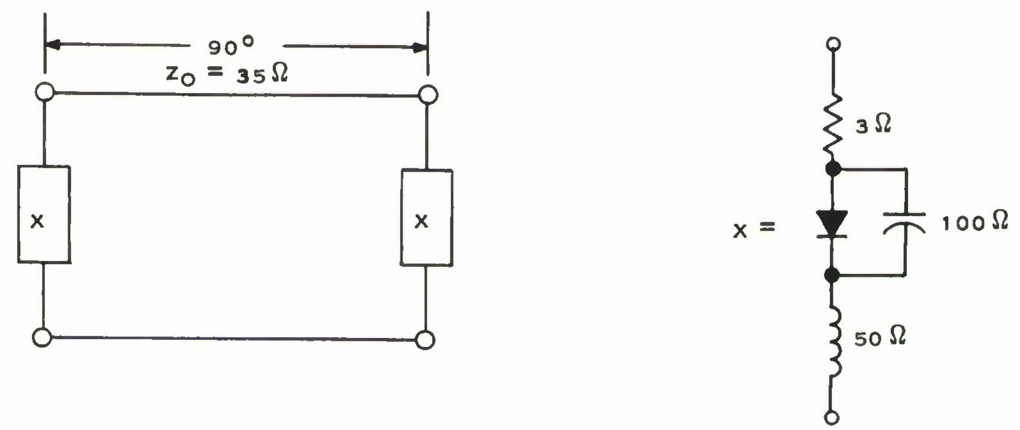
Figure 2. Schematic Diagram, 4-Bit Phase-Shift Network

A general closed-form solution for the insertion phase of a single phase-shift section is derived in Appendix I. Using the results from Appendix I the loading susceptances and line impedance are found to be $\pm j20.0$ millimhos and 35.3 ohms respectively. The $\pm j20.0$ millimho susceptances can be realized by using a diode with $-j$ 100-ohms reactance when reverse biased in series with a $+j$ 50-ohm reactance. Turning the diode off and on results in $\pm j$ 20.0-millimhos susceptance. A computer run was made for a single 90°-section and two 90°-sections in cascade. The results are shown in Figures 3 and 4, respectively. The losses are about



COMPUTED PERFORMANCE

NORMALIZED FREQUENCY	DIODE STATE	VSWR	LOSS dB	INSERTION PHASE	PHASE ERROR
0.9	ON	1.60	0.95	− 26.7	+50
0.9	OFF	1.66	0.53	−111.7	
1.0	ON	1.08	0.51	− 45.3	+0.5
1.0	OFF	1.08	0.51	−134.8	
1.1	ON	1.48	0.53	− 61.9	−18.7
1.1	OFF	2.64	2.30	−170.6	

PHASE ERROR $= 90^\circ - (\phi_{ON} - \phi_{OFF})$

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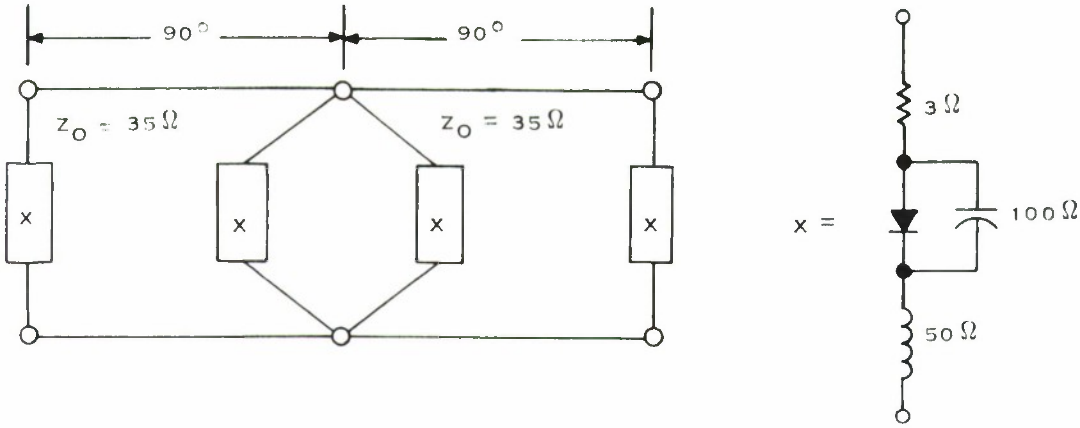
Figure 3. 90° Phase-Shift Section



COMPUTED PERFORMANCE

NORM-ALIZED FREQ.	DIODE STATE	VSWR	LOSS dB	INSERTION PHASE	PHASE ERROR
0.9	ON	2.16	2.33	− 52.2	+6.5
0.9	OFF	1.46	0.62	−225.7	
1.0	ON	1.11	1.02	− 90.6	+1.2
1.0	OFF	1.11	1.02	−269.4	
1.1	ON	1.44	0.82	−122.4	−37.3
1.1	OFF	4.9	6.20	−339.7	

PHASE ERROR = 180° − ($\phi_{ON} - \phi_{OFF}$)

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Figure 4. 180° Phase-Shift Section

what can be expected using 45°-sections, but the phase shift and VSWR, with respect to frequency, are worse.

Because manufacturing tolerances will cause a shift in the center frequency of the produced circuits it was decided to use 45°-sections as originally proposed. However, no circuits had been produced at the time effort on the program was stopped as a result of system redirection on Contract AF19(628)-5167.

B. UP-CONVERTOR

Frequency translation in the ESM module is performed by an upper side-band up-convertor. The input frequency band from 0.7 to 1.2 GHz is converted to an upper side band of 5.425 to 5.925 GHz with a pump frequency of 4.725 MHz. The theoretical conversion gain varies from 8.9 dB at 5.425 GHz to 6.94 dB at 5.925 GHz. In the proposal this was reduced to a 3-dB conversion gain across the band to account for circuit losses and the necessary fixed loss at the low end to remove the gain slope. The circuit was designed using micro-strip lines to realize the filter structures and thin-film resistors and capacitors for the bias circuit. This was in accordance with the desire for an all thin-film approach to meet the cost and reproducibility requirement in large-volume production.

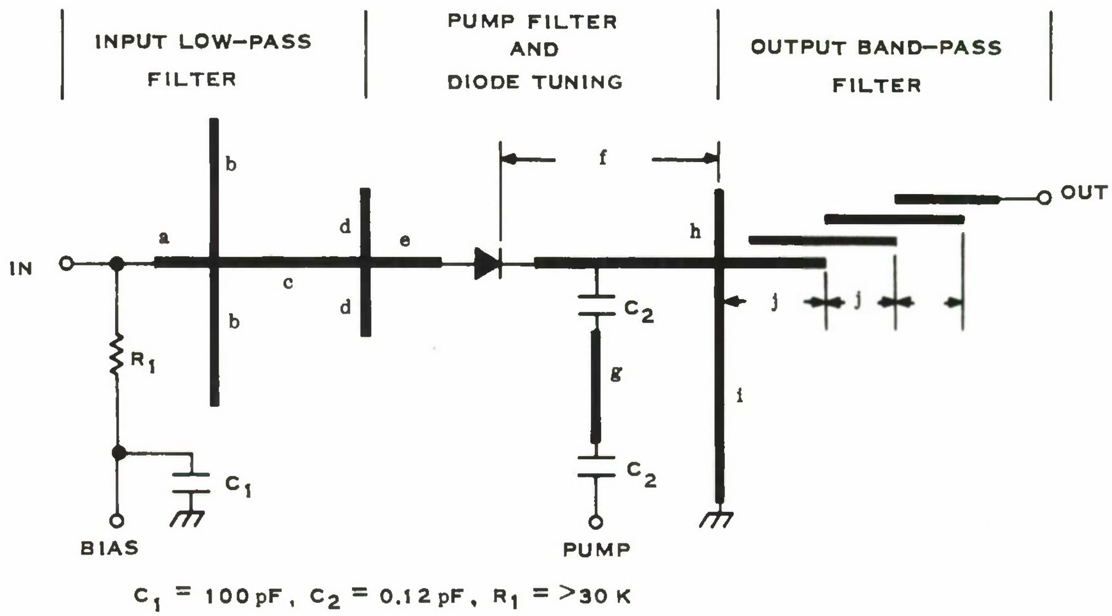
The design of wide-band up-convertors has been treated by several authors notably, Getsinger and Matthaei.[2,3] An attempt was made early in the program to synthesize the input and output filters following the method of the above authors. In essense, this consisted of designing the input and output filters as one composite band-pass filter which was then bisected and coupled together through the non-linear capacitance of the diode. The element values of the input and output filters are derived from the same low-pass prototype and then transformed to band-pass structures at the appropriate input- and output-frequency bands. This approach results in an efficient design using the minimum number of reactive elements.

However, difficulty is encountered in attempting to realize the above filter design for wide bandwidths using distributed elements especially μ-strip transmission lines. Another problem was the spurious pass bands of the input band-pass filter that must be kept away from the lower and upper side bands. These problems are not insurmountable but were considered beyond the

scope of this contract. Because of these difficulties a simpler approach was taken using a low-pass filter at the input. Although not as sophisticated as the above technique a working circuit could be designed and by optimizing element values with the aid of a computer the design objectives could be achieved.

The circuit consists of a low-pass filter at the input, a band-pass filter at the output and a varactor diode with the pump tuning circuitry connected in series between the input and output. In addition to series tuning the diode at input, output and the pump frequencies, the composite structure was designed to present a high-reactive impedance to the diode at the lower side band from 3.525 to 4.025 GHz. This is necessary to prevent any power flow at the lower side band which would result in undesired gain peaks in the conversion gain at the upper side band. A schematic diagram of the up-converter is shown in Figure 5. For convenience in the following discussion the various lines are lettered. Since the filter and diode tuning circuits are all interacting through the non-linear properties of the diode they will be discussed in the order that allows for a minimum iteration in the design approach.

A step-elastance diode was selected because it offers the possibility of constant conversion gain and circuit tuning independent of pump amplitude. This advantage is based on the fact that the step-elastance diode is switched between the S_{max} and S_{min} states at the pump frequency where S_{max} and S_{min} are fixed diode parameters and are independent of the pump amplitude. The duty cycle of the two states is a function of the pump amplitude and will effect circuit performance, but by a suitable choice of a self-bias resistor, this effect can be minimized. The characteristics of a step-elastance diode in an up convertor are derived in Appendix II. It is shown that a diode with S_{max} = $22.9 \times 10^{+11}$ darafs, S_{min} equals 0, and a 50% duty cycle provides a 50-ohm conversion impedance for the input and output



LINE	Z_0	LENGTH
a	80 Ω	0.203
b	33 Ω	0.262
c	80 Ω	0.492
d	33 Ω	0.128
e	50 Ω	0.133
f	50 Ω	0.480
g	50 Ω	0.405
h	47 Ω	0.082
i	47 Ω	0.492
j	Z_o = 37.5 Z_o = 87.5	0.200

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Figure 5. Up-Convertor Schematic

filters. This is a desirable feature due to the ease of the design and testing of the filters. A diode with these values is also readily available in a beam-lead form and will have virtually zero parasitic elements with which to contend.

The design of the input and output filters is straightforward because of the 50-ohm impedance level at their respective input and outputs. The input filter is a maximally flat, 4-element, low-pass design with a 3-dB cutoff at 1.8 GHz. A small area, low-pass design was chosen rather than a larger area band-pass design. Also, spurious pass bands are easier to control in a low-pass filter than in a band-pass type filter. The low-pass filter is comprised of lines, a, b, c and d in Figure 5. The output filter is of the maximally flat, band-pass type with a 3-dB band width of 900 MHz which is centered at 5.675 GHz. The band-pass filter is realized using quarter-wave, parallel-coupled lines (lines j in Figure 5). Its first spurious pass band is at 17.025 GHz and is outside the frequency band of interest. Both input and output filters were designed without grounded elements due to the difficulty of obtaining a good RF ground in a micro-strip medium.

Since a connection to ground is necessary on either side of the diode for bias purposes it was found convenient to place a shorted stub (line i Figure 5) at the input of the band-pass filter. This serves both as a ground return for the diode bias circuit and, by using a half wave length at the pump frequency, additional pump filtering at the output is provided. The open stub (line h Figure 5) provides a capacitive reactance to form a parallel resonant circuit with the inductive reactance of the shorted stub at the center of the output frequency band. Later, the impedance level of lines h and i will be selected to resonate the diode at the input frequency. The remaining dc return is provided by a thin-film resistor from the input of the low-pass filter to ground. The shunt impedance of this resistor, at the

input frequencies, is lower than the dc value due to the distributed capacitance of the thin-film pattern. An approximate solution for this shunt impedance can be found by solving for the characteristic impedance of a lossy transmission line. For a resistor pattern having 50 ohms per square and a 4-mil line width the impedance was calculated to be 700 -j 700 ohms which provides a negligible shunt loading at the 50-ohm input impedance of the low-pass filter.

Series tuning for the diode at the input, output, and pump frequencies is provided by lines e, f, h, and i together with the impedances at the output of the low-pass filter and the input of the band-pass filter. It is assumed that the loading effect of the pump-input filter (line g in Figure 5) is negligible because of its narrow bandwidth. The impedance level of line f is selected to be 50 ohms. Since the impedance at the input of the output band-pass filter (including the parallel-resonant circuit formed by lines h and i) is also 50 ohms, the length of line f will not affect the series resonance of the diode at the output frequency. Since the output impedance of the low-pass filter is essentially zero at the output frequency, it is only necessary to adjust the length of line e so that it forms a series-resonant circuit with the diode capacity at the output frequency. At the pump frequency the junction of lines h and i is a short circuit and the output impedance of the low-pass filter is essentially a short. Therefore, the lingth of line f can be adjusted to series resonate the diode capacity and line e at the pump frequency. Finally, the diode is series resonated at the input frequency by adding a half wave of 50-ohm line (at the pump frequency) to the length of f and by adjusting the impedance levels of lines h and i. Neither of these adjustments will affect the tuning of the diode at the pump and output frequencies.

The pump input filter (line g and capacitors C_1) is a narrow-band, half-wave resonator type with capacitive coupling at the

ends. The narrow band-width (4%) provides negligible loading on the circuit at the input and output frequencies. The pump filter was designed for 50-ohms input and output impedance. It is connected at a point in the diode series resonate circuit where the diode series resistance (~3 ohms) is transformed to 50 ohms.

This completes the first design of the up-convertor and was considered sufficient for constructing a development model. A circuit was constructed using the calculated line lengths and impedances shown in Figure 5. The dc return at the input of the low-pass filter was provided by a bias tee rather than a thin-film resistor to expedite the construction. A copy of the photo mask is shown as Figure 6.

The first circuit constructed was not successful. It was necessary to lower the pump frequency to approximately 4.5 GHz before energy could be coupled to the diode circuit. At this point, by using a panoramic receiver an upper side-band signal was observed at the output but it was much lower than the pump signal. The pump input filter was checked separately and found to be resonant at 4.5 GHz rather than 4.725 GHz. The half-wave shorted stub was also checked separately and found to be resonant at 4.675 GHz instead of 4.725 GHz. The error in the pump input filter was due to fringing capacitance at the ends of the line which had been neglected in the calculating of the line length. The error in the half-wave stub length was considered normal for the first trial circuit. Both lengths were adjusted and a new mask was ordered. The revised circuit had not been fabricated at the time system redirection occurred on Contract No. AF19(628)-5167.

The plans were to obtain test results from the first development circuit and do a redesign of the circuit using the computer to optimize the various line lengths.

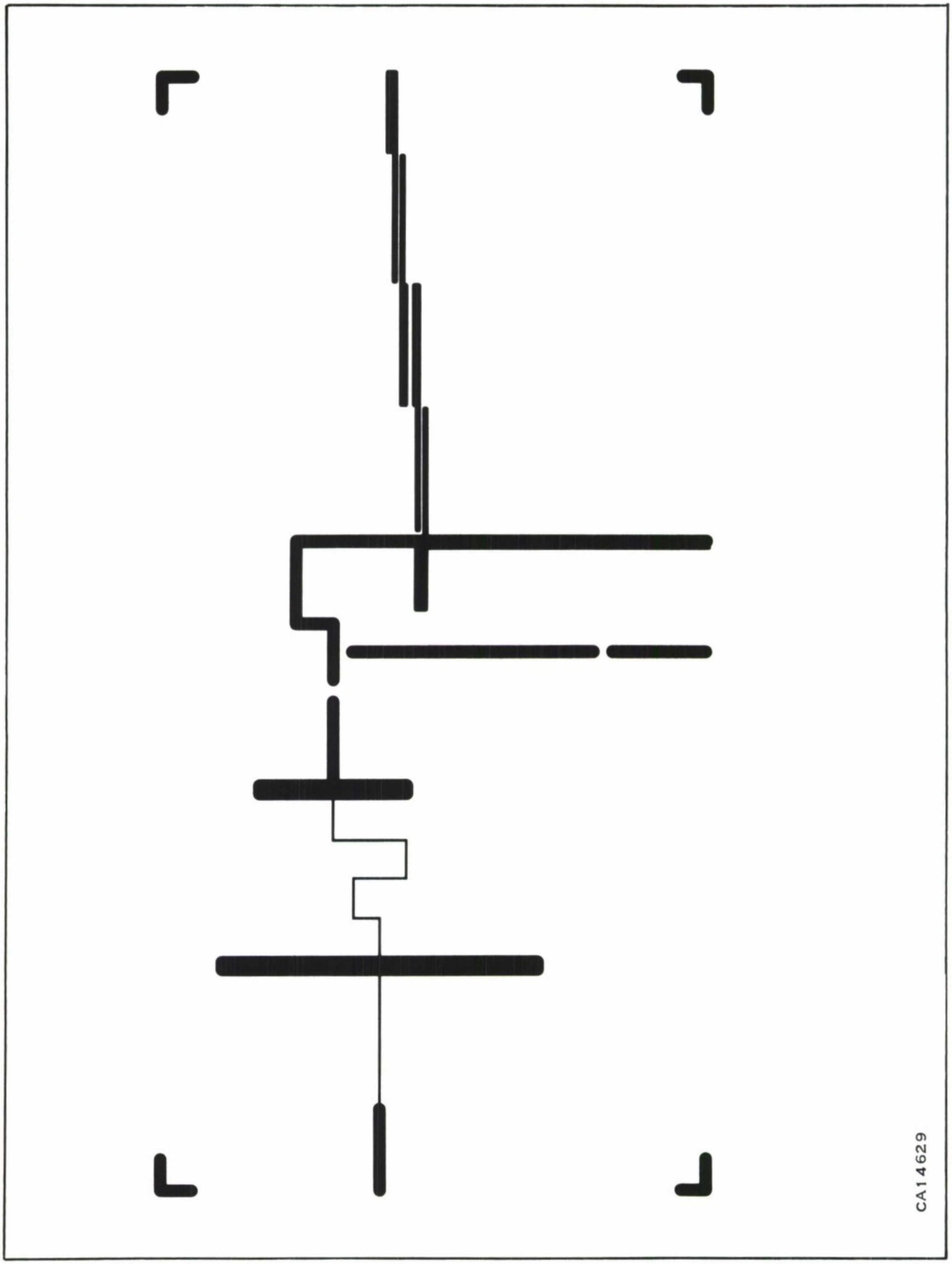

Figure 6. Up-Converter Mask

SECTION IV
CONCLUSION

At the time the system redirection occurred on Contract AF19(628)-5167 the design study had been completed and the first model of the steering module up-convertor had been fabricated and tested. Acceptable test data indicated changes necessary to meet module operational requirements. The revised circuit had not been fabricated at the time of redirection. The phase-shifter study was complete and initial layout was scheduled. Ancillary circuit functions posed no problem and hardware fabrication was planned.

In conclusion, Texas Instruments feels that the intent of this program was fulfilled. That is, the basic design approach was sound, fabrication of the steering module using thin-film techniques is both feasible and practical, and that the modules can be mass produced economically and reliably for use in large, ground-based phased-array radar systems.

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APPENDIX I

CLOSED-FORM SOLUTION FOR THE DESIGN PARAMETERS OF A SINGLE-SECTION LOADED-LINE-TYPE PHASE SHIFTER

APPENDIX I

CLOSED-FORM SOLUTION FOR THE DESIGN PARAMETERS OF A SINGLE-SECTION LOADED-LINE-TYPE PHASE SHIFTER

The relationship between a loaded-line phase-shifter section and an equivalent-unloaded uniform transmission line is shown below.*

$$\theta' = \cos^{-1}\left[\cos\theta - \frac{B}{Y_o}\sin\theta\right] \qquad (1)$$

$$Y_o' = Y_o\left[1 - \frac{B}{Y_o}^2 + 2\frac{B}{Y_o}\cot\theta\right]^{1/2} \qquad (2)$$

For digital phase-shifter applications the values of the loading susceptances are controlled in a discrete manner by the use of switching diodes. It is desirous to find a solution for the parameters of the loaded-line phase-shifter section given a desired change in the insertion phase with the contraint that $Y_o' = 1$.

Two special cases of interest are

1) The loading susceptances are switched from +B to -B, and
2) The loading susceptances are switched from zero (0) to some finite value B_1.

<u>Case 1</u>--Loading susceptances - ± B

Inserting the constraint $Y_o' = 1$ into Equation (2) we have

*White, J. F., High Power, PIN Diode Controlled, Microwave Transmission Phase Shifters, IEEE - PGMTT, March 1965, pp 233-242.

$$1 = Y_o \left[1 - \left(\frac{\pm B}{Y_o} \right)^2 + 2 \left(\frac{\pm B}{Y_o} \right) \cot \theta \right]^{1/2} \tag{3}$$

Equation (3) is satisfied for +B and -B only if $\cot \theta = 0°$ or $\theta = 90°$.

Letting $\theta = 90°$ and solving for B in Equation (3) we have

$$B = \sqrt{Y_o^{\,2} - 1} \tag{4}$$

substituting $\theta = 90°$, $B = +\sqrt{1 - Y_o^{\,2}}$ and $B = \sqrt{Y_o^{\,2} - 1}$ into Equation (1), we obtain the two following equations

$$\theta_+' = \cos^{-1} \left(- \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) \tag{5}$$

$$\theta_-' = \cos^{-1} \left(+ \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) \tag{6}$$

Since $Y_o' = 1$ for both +B and -B, the change in insertion phase is the difference between θ_+' and θ_-'. Let $\psi = \theta_-' - {}_+'$ be the desired change in insertion phase. Then

$$\psi = \cos^{-1} \left(+ \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) - \cos^{-1} \left(- \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) \tag{7}$$

$$\psi = \cos^{-1} \left(+ \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) - \left[\pi - \cos^{-1} \left(+ \sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) \right] \tag{8}$$

$$\psi = 2 \cos^{-1} \left(\sqrt{1 - \frac{1}{Y_o^{\,2}}} \right) - \pi \tag{9}$$

$$\frac{\psi}{2} + \frac{\pi}{2} = \cos^{-1}\left(\sqrt{1 - \frac{1}{Y_o^{\,2}}}\right) \tag{10}$$

$$\cos\left(\frac{\psi}{2} + \frac{\pi}{2}\right) = \sqrt{1 - \frac{1}{Y_o^{\,2}}} \tag{11}$$

$$-\sin\frac{\psi}{2} = \sqrt{1 - \frac{1}{Y_o^{\,2}}} \tag{12}$$

$$\sin^2\frac{\psi}{2} = 1 - \frac{1}{Y_o^{\,2}} \tag{13}$$

$$\frac{1}{Y_o^{\,2}} = 1 - \sin^2\frac{\psi}{2} = \cos^2\frac{\psi}{2} \tag{14}$$

$$Y_o = \frac{1}{\cos\frac{\psi}{2}} = \sec\frac{\psi}{2} \tag{15}$$

Substituting $Y_o = \sec\frac{\psi}{2}$ into Equation (2) we have

$$B = \sqrt{\sec^2\frac{\psi}{2} - 1} \tag{16}$$

$$B = \sqrt{\mathrm{Tan}^2\frac{\psi}{2}} = \mathrm{Tan}\frac{\psi}{2} \tag{17}$$

Thus, for the case where the loading susceptances are switched between $+B$ and $-B$ the parameters are, $\theta = 90°$, $B = \pm\,\mathrm{Tan}\ \psi/2$ and $T_o = \sec\psi/2$, where ψ is the desired insertion phase.

Case 2--Loading susceptances = 0 and B_1

$$1 = Y_o \left[1 - \left(\frac{B}{Y_o} \right)^2 + 2 \frac{B}{Y_o} \cot \theta \right]^{1/2} \tag{18}$$

Equation (18) is satisfied for both $B = 0$ and $B = B_1$ only if $Y_o = 1$. Letting $Y_o = 1$ and $B = B_1$ in Equation (18), and by squaring both sides, we have

$$1 = 1 - B_1^2 + 2 B_1 \cot \theta \tag{19}$$

$$B_1 = 2 \cot \theta \tag{20}$$

Substituting $Y_o = 1$, $B - 0$ and $B = 2 \cot \theta$ into Equation (1), we obtain the following two equations

$$\theta_0' = \cos^{-1} (\cos \theta) = \theta$$

$$\theta_{B_1}' = \cos^{-1} (\cos \theta - 2 \cot \theta \sin \theta) \tag{21}$$

$$\theta_{B_1}' = \cos^{-1} (-\cos \theta) = \pi - \theta \tag{22}$$

Let $\psi = \theta_0' - \theta_{B_1}'$ be the desired change in insertion phase as in Case 1. Then

$$\psi = \theta - (\pi - \theta) = 2 \theta - \pi \tag{23}$$

$$\theta = \frac{\psi}{2} + \frac{\pi}{2} \tag{24}$$

Substituting the value for θ in Equation (24) into Equation (20) we have

$$B_1 = 2 \cot \frac{\psi}{2} + \frac{\pi}{2} \tag{25}$$

$$B_1 = -2 \operatorname{Tan} \frac{\psi}{2} \tag{26}$$

Thus, for the case where the loading susceptances are switched between zero and B_1 the parameters are $Y_o = 1$, $B_1 = -2 \tan \psi/2$ and $\theta = \psi/2 + \pi/2$, where ψ is the desired change in insertion phase.

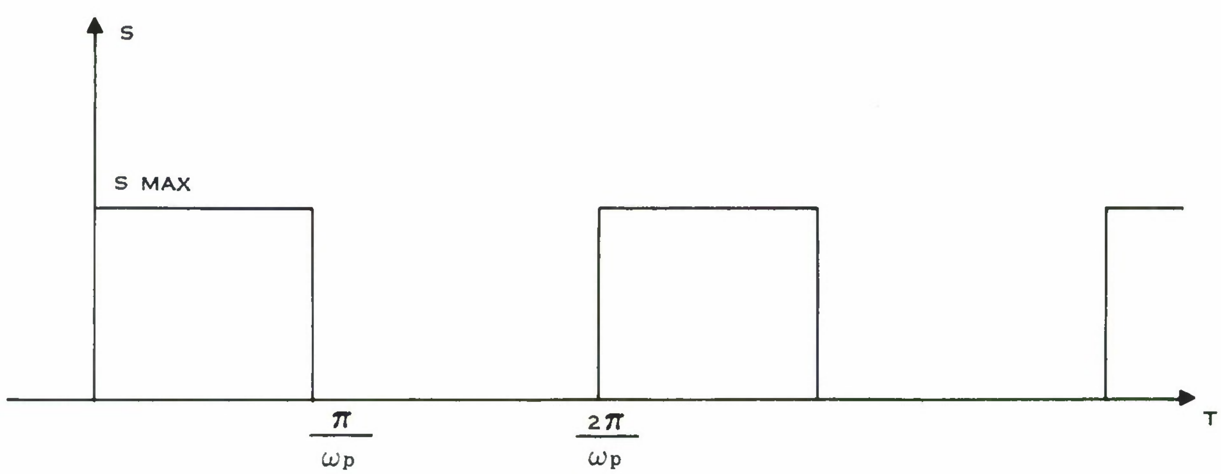
APPENDIX II

ANALYSIS OF THE STEP-ELASTANCE DIODE FOR USE IN UP-CONVERTOR

APPENDIX II

ANALYSIS OF THE STEP ELASTANCE DIODE FOR USE IN AN UP-CONVERTOR

The step-elastance diode for the purpose of this analysis is assumed to have an elastance $S = 1/c$ of zero when the diode is in the conducting state; and a constant elastance of S_{max} when the diode is turned off. The diode series resistance is assumed negligible. Beam-lead diodes are available which are very good approximations of the above. We will also assume that the diode is pumped so that it changes state each one-half cycle of the pump frequency producing the following elastance V_s time curve. The



elastance, as pumped, can be written in Fourier series

$$S\ (t) = S_0 + \sum_{k=-\infty}^{\infty} S_k\ e^{\ jkwpt}$$

where

$$S_k = \frac{S_{max}}{k\Pi}\ , \text{ when k is odd}$$

$$S_k = 0, \text{ when k is even}$$

$$S_0 = \frac{S_{max}}{z}$$

If the diode is open-circuited at all frequencies except the input, pump and upper sideband frequencies, we can write the small signal equations as follows

$$\begin{bmatrix} V_s \\ \\ V_u \end{bmatrix} = \begin{bmatrix} \frac{S_o}{j\omega_s} & \frac{S_1^*}{j\omega_u} \\ \\ \frac{S_1}{j\omega_s} & \frac{S_o}{j\omega_u} \end{bmatrix} \begin{bmatrix} I_s \\ \\ I_u \end{bmatrix}$$

The notation follows that of Penfield and Rafuse,† where the subscripts s and u refer to the input and upper sideband frequencies respectively. The optimum source and load resistance for maximum gain and minimum noise figure has been solved using the above equations and is as follows

$$R_{opt} = \sqrt{\frac{S_1\ S_1^*}{\omega_s\ \omega_u}}$$

It is convenient in designing the input and output filters to work at an impedance level of 50 ohms. For an input center frequency of 0.95 GHz and an output center frequency of 5.675 GHz the value of $S_{max} = \pi S_1$ is calculated to be 22.9×10^{-11}. This corresponds to a diode with a reverse bias capacity of 0.536 pF. The average capacity under pumping conditions will be twice this

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†Penfield, P., and Rafuse, R. P., "Varactor Applications," *The MIT Press*, Cambridge, Mass. (1962).

value or 0.872 pF. This average capacitance must be incorporated in the design of the input, output, and pump filter circuits.

It is interesting to note that the elastance wave form produced by the pump signal can also be generated by periodically opening and closing a switch connected across a capacitor. This is not a valid equivalent circuit for the pumped-step elastance diode in an up-convertor circuit since the Manley-Rowe power formulas require the power at the pump and signal frequencies interact in the nonlinear reactance. If we applied a signal to such a switched capacitor we would expect sidebands to be generated, but there should be no conversion gain.

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13. ABSTRACT

This is the final technical documentary report for the Electronic Steering Module program, Contract No. C-595, prepared for Lincoln Laboratory, Massachusetts Institute of Technology. This report documents all efforts expended from the contractual start date, 17 April 1967, to 15 January 1968. The overall goal of this program was to design an element steering module that would be easily produced in large quantities at a minimum cost. The effort of this program was directed toward the design of the ESM in a microstrip form utilizing thin-film techniques to form the resistors, capacitors and transmission-time elements on a ceramic substrate.

14. KEY WORDS

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